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L. EDELEN
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INVENTOR(S)/APPLICANT(S)			
LAST NAME	FIRST NAME	Middle Initial	RESIDENCE (CITY AND EITHER STATE OR FOREIGN COUNTRY)
Hoult	David	Ian	28 Smithfield Ave., Winnipeg, Manitoba, R2V 0B6, Canada

TITLE OF THE INVENTION (280 CHARACTERS MAX.)

Cartesian Feedback for Precise Quantification and Superior Performance of Magnetic Resonance Instruments

CORRESPONDENCE ADDRESS

J. Wayne Anderson
National Research Council of Canada
Intellectual Property Services Office, EG-10, Bldg. M-58
Montreal Road, Ottawa, Ontario, Canada K1A 0R6

STATE	Ontario	ZIP CODE	K1A 0R6	COUNTRY	Canada

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Respectfully submitted,

J. Wayne Anderson
Patent Agent for Applicant
Regn No. 28,158

Enclosures
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___ Additional inventors are being named a separately numbered sheets attached hereto.

PROVISIONAL APPLICATION FILING ONLY

Cartesian Feedback for Precise Quantitation and Superior Performance Of Magnetic Resonance Instruments

BACKGROUND OF THE INVENTION

[01] There are numerous defects and deficiencies of magnetic resonance instrumentation to which the user becomes first resigned then oblivious. Perhaps chief among these is that a spectrometer or imager is uncalibrated in an absolute manner. Thus when transmitting, a change in or of the sample (e.g. non-conducting to conducting, polar to non-polar, etc.) necessitates a change of pulse power or length, even if the probe has been re-tuned and re-matched – and the latter may not be possible if automatic sample-changing is employed. Concomitantly, upon reception, even if different samples contain the same number of nuclei, the amplitudes of their free induction decays (FID) following a 90^{0} pulse (i.e. the integral of the spectrum) may differ. With an ideal instrument, such changes would not occur. The user would dictate the type of pulse, its flip-angle and bandwidth or duration, and the settings would then remain fixed regardless of the sample. Equally, for constant sample temperature, the amplitude of the FID would depend solely on the receiver gain and the number of nuclei (we discount NMR 'invisible' nuclei), once again regardless of the sample composition. Thus, to take but two examples among many, there would be no need to worry that the swelling of a perfused heart would change the pulse length and the signal strength of some metabolite of interest, or that the breathing of a patient in an imaging experiment would do likewise.

[02] These shortcomings are caused by the tuned and matched probe, for its Q-factor and tuning are dependent on its environment; in particular, both variables are sample-dependent. A further related defect is that tuning can be modulated by movement and vibration, the latter causing extra "1/*f*" noise about a resonance and "t1" noise in two dimensional spectra via phase modulation. Such vibration can come from air flow, from spinning the sample, from gradient noise and even from loud music played near the instrument! Spinning the sample in high resolution or magic angle experiments can also directly modulate the probe tuning if there is any heterogeneity or asymmetry in the electric susceptibility of the sample. This causes extra spinning sidebands. If more than one tuned probe coil at the same frequency is used, these sensitivities can increase greatly due to coupling. Examples are high resolution quadrature probes, coils in rotating frame imaging experiments, phased array coils, etc. While "paddles" [1] or capacitor-resistor leakage bridges [1,2] can cancel both reactive and resistive interactions, the sensitivity of the balance still renders such solutions very susceptible to all the factors mentioned above. To cap this litany of woes, change of temperature also affects probe tuning and Q-factor, causing B_1 field and signal strength to drift in amplitude and phase.

[03] Now during signal reception, the use of pre-amplifiers with a large voltage standing wave ratio (VSWR) has been known for many years to reduce the *effective* probe Q-factor, without affecting signal-to-noise ratio [3-6], thereby ameliorating the problems. Nevertheless, this technique is still neither widely used nor understood. During transmission, the equivalent is to over-couple the transmitter [3], but gains in this regard come at great expense, as the transmitter power specification has to be seriously

increased for a useful effect to be obtained; further, maintaining a constant degree of over-coupling is difficult with systems designed for only 50 Ω use. Thus problems during transmission remain unresolved.

[04] Turning to other ills, when the signal is very strong, "radiation" damping can be a nuisance. It is usually partially dealt with by detuning the probe and/or reducing the flip angle, though another solution pertinent to this paper exists [7]. The use of a large VSWR pre-amplifier can again help in this regard [8], its beneficial effect first being observed at the Perkin-Elmer Corporation in Beaconsfield U.K. in about 1970, but in some cases this solution may be inadequate. Next, crossed diodes in probe protection circuits cause considerable non-linearity at low pulse powers and as a result, can greatly distort selective pulses. More expensive PIN diode circuits help considerably here, but even they are not distortion-free and their switching times are often excessive. Finally, high power transmitters are renowned for changes of power and phase during a pulse as components rapidly heat and power supply voltages droop. Pulses are then distorted in amplitude and phase and of course, ambient temperature changes have their destructive role to play here too. Distorted selective pulses might, for example, in a biological experiment cause water to be inadvertently excited, leading to unacceptable water signal suppression. However, our purpose in this Communication is not merely to detail defects, but to describe a solution to all these ills.

Negative Feedback

[05] Negative feedback has been renowned since the pioneering work of Black [9] for its ability greatly to reduce amplifier distortion. Essentially, it functions by sampling the output of an amplifier, comparing it to that desired and correcting the output to cancel any error. Thus, in a magnetic resonance context, an example of its use would be during transmission. With a small sense loop, or by monitoring a small fraction of the voltage across a tuning capacitor, we could directly or indirectly sample the B_1 field produced by a probe and compare the monitored signal with that desired, as shown in Fig. 1a. With sufficient open-loop gain, any distortion introduced by the power amplifier or by crossed diodes would then be reduced to negligible levels. Ignoring for the moment the massive practical difficulties to be overcome in attempting such a design, the simple equivalent on-resonance circuit of Fig. 1b is the basis for an on-resonance calculation. Initial tuning and matching effectively reduces the transmitter source resistance Z_0 to a resistance r_0 equal to the probe resistance r at that time. Concomitantly, by conservation of energy, the transmitter EMF is reduced by a factor $\eta = \sqrt{r_0 / Z_0}$. Compensating for phase shifts (not shown), we then have that the open-loop sense coil voltage V_s is given by

$$V_s = i_p \omega_0 M = \omega_0 M \left(\frac{\alpha \eta V_{in}}{r + r_0} \right) \equiv \alpha \eta \beta V_{in}; \quad \beta = \frac{\omega_0 M}{r + r_0} \qquad (1)$$

where α is the gain of the transmitter, V_{in} is its input voltage, M is the mutual inductance between the probe and the sense loop, i_p is the current in the probe coil and ω_0 is the angular Larmor frequency. The factor β is essentially the attenuation introduced by the small mutual inductance. Assuming any losses introduced by the sense coil are

negligible.) If the switch is now thrown and the feedback loop closed, we obtain the classic feedback formula

$$V_s = \alpha\eta\beta(V_{in} - V_s) \qquad (2)$$

and hence

$$V_s = \frac{\alpha\eta\beta}{1+\alpha\eta\beta} V_{in} \qquad (3)$$

[06] The probe current i_p is then $V_s/\omega_0 M$. Clearly, if the open-loop gain $\alpha\eta\beta >> 1$, the probe current has negligible dependence on transmitter gain α, ? ? ? ? ? ? ? ? η and probe resistance r, and is to all intents and purposes $V_{in}/\omega_0 M$. In other words, the B_1 field, which, of course, is proportional to probe current i_p, is essentially independent in amplitude and phase of the vagaries of the system – an enormous asset. Any distortion introduced by the transmitter or crossed diodes (essentially a variation of α) is also rendered negligible – the well-known attribute of negative feedback. A more thorough analysis at all frequencies shows that changes of tuning are also irrelevant within wide limits. The system has become insensitive to probe perturbations caused by vibration, spinning, change of sample, change of coupling, etc. The B_1 field remains what it should be and a probe can be calibrated absolutely. All that is needed is knowledge of the mutual inductance M and of the relationship between the probe current and the B_1 field in the rotating frame. In essence, the negative feedback circuit demands that the voltage V_s induced in the sense coil be always the same as the input voltage V_{in}. As the probe loading, tuning or matching change, the output of the transmitter automatically compensates. Only if the error is so great that the output of the transmitter attempts to exceed its maximum power does the scheme fail.

[07] Another instructive way of looking at the functioning of the circuit is to rewrite Eq. (3) in the form

$$i_{\mathrm{p}} = \frac{V_s}{\omega_0 M} = V_{\mathrm{in}} \frac{\alpha\eta\beta}{\omega_0 M(1+\alpha\eta\beta)} = V_{\mathrm{in}} \frac{\alpha\eta}{(r + r_0 + \alpha\eta\omega_0 M)} \tag{4}$$

[08] If we initially have $r = r_0$, the quantity $G = \alpha\eta\omega_0 M/2r_0$ is the forward loop gain which we have set to a large value, say 100. Then Eq. (4) becomes

$$i_{\mathrm{p}} = V_{\mathrm{in}} \frac{\alpha\eta}{r + (1+2G)r_0} \tag{5}$$

[09] The effective transmitter source resistance r_0 has been greatly increased by the order of the loop gain, and the transmitter has essentially become a constant current source without resorting to gross over-coupling and without the expenditure of extra power. This facet of feedback can be exploited to the hilt when multiple transmitter coils are desired [6] for surrounding coils do not then influence the current flow, and this is the subject of the next Communication [10].

[010] Turning to signal reception, the roles of the matching and the sense coil are reversed, as shown in Fig. 1c. However, as in general the input impedance r_{p} of the pre-amplifier is not Z_0 (50 Ω), its transformed resistance R_{p} (Fig. 1d) is not initially r. Indeed, R_{p} may by design [3-6] be much greater than r without loss of signal-to-noise ratio. With reference to Figs. 1c and 1d, we may again write an equation for the open-loop gain. If the induced EMF of a free induction decay is ξ, the output voltage of the pre-amplifier is

$$V_{\mathrm{out}} = -\alpha \frac{R_p \xi}{r + R_p} \equiv -\alpha\gamma\xi; \quad \gamma = \frac{R_p}{r + R_p} \tag{6}$$

[011] However, when the loop is closed, a second voltage V_s is induced in the receiving coil by the current in the sense coil and the probe is the summing point of the circuit. Assuming the reactance of the sense coil $\omega_0 M \ll R_s$, this voltage is approximately $\omega_0 M V_{out}/R_s$, and so, from Eq. (6) we have the feedback equation

$$V_{\text{out}} = \frac{-\alpha\gamma\xi}{1+\alpha\gamma\omega_0 M / R_s} \tag{7}$$

[012] If $\alpha\gamma\omega_0 M/R_s \gg 1$, the resultant has negligible dependence on α and γ. The output voltage is then

$$V_{\text{out}} \cong -\xi \frac{R_s}{\omega_0 M} \tag{8}$$

[013] In other words, the output signal is essentially independent of the details of the probe tuning and matching. The negative feedback is striving to keep to zero the input voltage of the pre-amplifier and so the voltage induced by the current in the sense coil almost cancels the induced EMF. As the output voltage no longer depends on the vagaries of the probe, but rather upon the resistance R_s and the mutual inductance of the receiving and sense coils (both under the designer's control), it follows that the probe can be permanently calibrated and a known number of nuclei always induces an FID of constant amplitude virtually regardless of probe Q-factor or tuning.

[014] Just as with the transmitter the negative feedback was considered to have greatly increased the former's *output* impedance, so now the effective *input* impedance R_p of the pre-amplifier is greatly increased. From Eq. (7), setting $V_{out} = -\alpha i_p R_p$, where i_p is the current flowing in the probe and into the pre-amplifier, we have that

$$i_{\mathrm{p}} = \frac{\xi}{r + R_{\mathrm{p}}\left(1 + \alpha\omega_0 M / R_{\mathrm{s}}\right)} \qquad (9)$$

[015] The current in the probe created by the FID is greatly reduced by the order of the loop gain, which has implications for "radiation" damping and for coupling to another tuned circuit nearby - both are greatly lessened.

Cartesian Feedback

[016] The impracticality of the above analysis must be immediately acknowledged. Any attempt to produce feedback of the type described will inevitably result in oscillation, for the Bode feedback stability criteria [11] are impossible to fulfil. Even if all the electronics associated with a design could have so broad a bandwidth that the group delay were negligible, the length of cable needed between the transmitter and probe would nullify the attempt. For example, with a 3 T (128 MHz) imaging magnet, there might be a run of 10 m of cable from transmitter to probe and the same run back from the sense loop. With average coaxial cable, the delay introduced would then be about 100 ns. With a probe having a Q-factor of 100, a 180° phase change would be reached only 2.85 MHz off resonance and the feedback would be positive, making the circuit ripe for oscillation. Indeed, an open-loop gain of less than only 4.56 would then be demanded for stability. The improvement in instrument performance with such a small gain is hardly worth having.

[017] A solution to this problem is to restrict severely the bandwidth with the aid of a sharp filter so that the open-loop transfer function is dominated by the filter characteristics. A typical NMR spectral width is, after all, usually much less than the

bandwidth $\omega_0/2\pi Q$ of the probe. If we take a "worst case" scenario of a probe Q-factor of 500 and a group delay of 300 ns (we have included delays in the electronics), a filter at 128 MHz having a Q-factor of 36,000 (bandwidth 3.56 kHz) can deliver an open-loop gain of up to approximately 50 dB (316) before oscillation commences. In other words, a maximum gain-bandwidth product of ±560 kHz is possible without oscillation. For those unfamiliar with the details of feedback, it should not be thought that the application of such a narrow bandwidth restricts the spectral width to 3.56 kHz. On the contrary, a bandwidth of over 1 MHz is possible with our example *once the loop is closed*. Rather, the *advantages* of the feedback (reduction of distortion, etc.) are only fully realised over the filter bandwidth and slowly diminish as the open-loop gain decreases away from the filter's centre-frequency. The problem now becomes that of how to make the desired filter, for it has an enormous Q-factor yet must be highly stable.

[018] A limited solution for transmission only was published by Hutchison *et al.* in 1980 [12]. With this "envelope feedback" method, the pulse is rectified in a linear detector to obtain its magnitude, and the filtered resultant is then used for comparison with the desired pulse waveform. However, this technique is single quadrant and therefore of no use if the pulse is bi-polar, as are most selective pulses. Meanwhile, in the communications industry, the solution now known as Cartesian feedback was proposed by Petrovic in 1983. [13]. (Petrovic also proposed so-called "polar feedback" methods, but it is not obvious that these techniques can maintain optimum signal-to-noise ratio.) Independently, the same full four-quadrant solution in the context of NMR was proposed by DIH in 1989 [14] and described by him in 1991 [15]. Dealing with transmission only

at the time, he realised that a high-Q filter could be made by quadrature (I-Q) detection of the signal from the sense coil, followed by resistor-capacitor (RC) filtering of the two quadrature channels, followed by I-Q re-modulation. It is, of course essential that the phase of the detection and/or re-modulation be such that *negative* feedback is applied when the loop is closed. An incorrect phase could result in oscillation. The criticism is unfounded, as shown independently in 1995 by Broekaert and Jeneer [7] who reported using Cartesian feedback to reduce greatly "radiation" damping.

SUMMARY OF THE INVENTION

[019] According to one embodiment of the invention we have successfully implemented a full Cartesian feedback spectrometer where transmitter and receiver are on at all times and where electronic switches route signals and feedback to and from the probe and sense coil appropriately. The instrument currently functions at 128 MHz (1H @ 3 T). In another embodiment of the invention, Cartesian feedback has been successfully applied to a phased array of coils for the purpose of transmission in magnetic resonance imaging.

DETAILED DESCRIPTION OF THE INVENTION

A "Hi-fi" Spectrometer

[020] According to one embodiment of the invention, two instruments that intersect at the probe (transmitter and feedback ⇒probe ⇒ receiver and feedback) are provided.

[021] According to a second embodiment, the receiver is used for feedback when transmitting, and then use the transmitter for feedback when receiving, albeit with the omission of the power amplifier. Both methods have their advantages and drawbacks

[022] The second method is shown in Fig. 2.

[023] *Probe and Sense Coil:* We note that a sense coil is but one method of detecting the current in the probe and there may be circumstances where it is better to detect the conservative electric field at the tuning capacitor(s) [10].

[024] *Modulation and Demodulation:* As digitised feedback systems can oscillate with 1-bit amplitude, the instrument is perforce analogue at this time. The matching of the two (I and Q) RC low-pass filters should dominate the closed-loop performance, and the orthonormality of the quadrature detection and re-modulation processes. A specialised unit [16] working at an intermediate frequency of 10 MHz with phase accuracy of 0.1° and amplitude accuracy of 0.1%, was therefore employed. Note that once the loop is closed, any quadrature-phase errors propagate round the loop. The choice of intermediate frequency is a trade-off between conflicting requirements: a higher frequency eases filtering problems that cause group delay (*q.v.*) but this is offset by the need for faster components restricted to military designs, by higher levels of crosstalk between switching signals and by greater difficulty in filtering switching signal feedthrough.

[025] *D.C. Offsets:* Direct voltage offset on the FID is usually considered unimportant as it can be removed during data analysis. However, with a feedback spectrometer, voltage offset during reception translates into a constant small signal at the Larmor frequency's being injected into the probe, which could cause nuclear saturation. Attention must therefore be paid to offsets caused by operational amplifiers in the demodulator, modulator and attendant filters. Achievement of minimum direct offset and flicker noise conjointly with maximum gain-bandwidth product involves conflicting constraints and innovative approaches such as the use of composite wideband and chopper-stabilised amplifiers. In particular, many wideband amplifiers tend to have high power deposition which increases the likelihood of thermal drift.

[026] *Phase:* The solution of the "phase problem" mentioned above is trivial, for both loop gain and phase can be monitored at any time by a computer and adjusted accordingly. However, widely different phases and gains are needed during transmission and reception. It is therefore necessary to be able to switch quickly between two phases and to this end, we employ three synchronised phase-locked loop (PLL) 80 MHz oscillators that are divided down in frequency. One fixed-phase oscillator is the reference for the modulator, while the other two variable-phase oscillators are references for the receiver quadrature demodulator. We do not attempt to alter the phase of an individual oscillator when changing from transmission to reception and *vice versa*; rather, we switch oscillators as this is much faster. Correct initialisation and synchronisation of the three oscillators at "power-up" is critical and they must be utterly immune to power transients. They must also operate smoothly over a full 360° range.

[027] *Gain*: Gain is a complex issue (in both senses), and it is important to keep keenly in mind that it is *feedback* that determines closed-loop gain. Thus, when receiving signal, it is the transfer function of the *transmitter* chain (modulator⇒IF stage⇒ mixer ⇒sense coil) that is important. Conversely, when transmitting, it is the transfer function of the *receiver* chain (sense coil⇒mixer⇒IF stage⇒demodulator⇒AF stage) that is important. Once the instrument is under computer control, such confusing detail can be obscured and less bemusing language devised. When setting up for transmission, we typically first work in traditional (open-loop) mode. With a simple "pulse and acquire" sequence, the pulse is generated by injecting, as usual, an appropriate voltage (say 2 V) into the real port of the transmitter modulator for the requisite period (e.g. 40 μs) and setting the transmitter gain (modulator⇒IF stage⇒ mixer ⇒power amplifier⇒probe) in the IF stage. However, we then go a step further, and monitoring the resulting B_1 field with the sense coil and the receiver, set the receiver gain and phase so that the quadrature AF output is the same complex voltage: 2 V real. In other words, the open-loop gain is set to complex unity, *to the resolution available in the receiver gain circuitry.* This ensures that when feedback is applied later, the flip angle will remain approximately the same. From this point on, the *receiver* settings pertinent to transmission are not touched, for they determine the calibration of the transmitter.

[028] Equally, when receiving signal, we initially set the receiver gain to give a reasonably-sized FID (say ~ 2V). We then turn the pulses off, and continuously inject a small signal at the Larmor frequency into the probe using the transmitter curcuitry: modulator⇒IF stage⇒ mixer ⇒sense coil. The transmitter IF gain (in practice, usually

attenuation) is set to give, *once again within the limits of the gain resolution*, unity open-loop gain. From this point on, these *transmitter* settings pertinent to FID signal reception are not touched, for they determine the calibration of the receiver. We then reduce the audio bandwidth by introducing the matched RC low-pass filters, turn on the feedback and increase the forward loop gains by 40 dB. In other words, during transmission when an RF pulse is generated, the transmitter gain is increased by 40 dB, while during FID signal reception, the receiver gain is increased by 40 dB.

[029] It follows that the dynamic range of the gain in both the receiver and transmitter chains must be sufficient to accommodate this procedure. On the face of it, it appears that there could be gross overload (*q.v.*) of the IF stages; however, it must be remembered that feedback is being applied and so this is not the case. There is large scope here for computer control and characterisation of the instrument. For example, if the gains of the instrument are precise and the characteristics of the sense coil are known, absolute calibration of flip angle and signal strength are possible. We ultimately envisage probes/sense coils having electronic coding that informs the computer of the probe being used. Probe characteristics are then retrieved from memory. If the gain resolution is only say 1 dB, the computer can compensate by varying to a minor extent the voltage into the modulator (from quadrature digital-to-analogue converters) during transmission and the gain of the digitised signal during reception.

[030] *Overload*: To minimise the group delay (*q.v.*), the bandwidths of both receiver and transmitter have to be large – say ±10 MHz about 128 MHz. Now outside the closed-

loop bandwidth (e.g. ±200 kHz), the receiver chain has its full gain up to the audio RC filters. There is therefore the risk, during reception with small signal strengths and at high gain, of overloading the final quadrature detector with out-of-band noise that is never seen by the analogue-to-digital converters. Overload detection at the intermediate frequency immediately prior to demodulation is therefore advantageous. Equally, an overload detector in the transmitter chain is useful, as it can sense if the transmitter cannot apply sufficient power. This can happen if a change in conditions lowers the Q-factor of the probe, and the negative feedback, in attempting to compensate, tries to apply more power than the transmitter can provide.

[031] *Noise*: It is an interesting point that if Cartesian feedback preserves the amplitude of the FID, the noise floor must rise if the Q-factor of the probe drops with change of sample. This is the exact opposite of the usual behaviour, where the probe is re-matched to 50 Ω with change of sample and the signal then drops while the noise remains constant. Note that the feedback neither decreases nor increases the signal-to-noise ratio (S/N) unless there is an engineering error.

[032] *Group Delay*: With multiple stages of amplification and/or attenuation in both transmitter and receiver at radio and audio frequencies, group delay can increase frighteningly quickly. Particular attention must be paid to the various filters needed and the bandwidth of the audio amplifiers. Our demodulator, like others, responds to odd harmonics and so elliptic filters with zeros at 30 MHz and 50 MHz were employed. We have been able to reduce the group delay through the electronics to about 250 ns, to

which must be added the line lengths to the probe and the transfer function of the probe itself. It is worth noting that, at least in theory, compensation for the delay can be introduced in the intermediate frequency sections of the receiver and transmitter chains. In this regard the use of single-pole (RC) I and Q audio filters having a cut-off frequency that is half the greatest bandwidth (say 50 kHz), while still allowing an open-loop gain of 40 dB to be applied is contemplated. Then the advantages of feedback are equally applied over that bandwidth.

[033] *Gain Distribution and Variable Gain Control:* An inherent problem with the Cartesian feedback method is that while any distortion originating in the forward path is greatly reduced by the feedback, any distortion that arises in the feedback path itself undergoes no reduction. Care must therefore by taken with dynamic range issues and in particular, attention was paid to linearity in the design of the quadrature modulator and demodulator. With regard to the receiver IF amplification chain, true variable gain amplifiers under digital control are, as usual, preferable to maximise dynamic range. Regrettably, most commercial "variable gain" amplifiers are actually fixed gain with variable attenuation.

[034] *Dynamic Stability:* Disturbances to loop stability are inherently associated with the process of switching from transmit to receive and *vice versa*. In particular, as shown in Fig. 2, the position of the low-pass RC filtering in the feedback loop has to be moved to maintain optimum gain bandwidth and stability, and to minimise transient generation. A considerable stability margin must be maintained to ensure "well-behaved" (no

ringing, overshoot, etc.) closed-loop operation and we typically remain 45° away from the 180 ° instability point. This reduces the gain-bandwidth product and we routinely work with a 1 kHz RC filter (2 kHz RF bandwidth) and 40 dB (x100) gain. This also has the advantage of introducing a considerable margin of safety in case the transmitter and/or probe characteristics alter significantly with temperature and time. In this regard, while the computer should monitor open-loop transfer function periodically and adjust it accordingly, it is always a good idea to have fail-safe protection such as an RF fuse on the output of the power amplifier! Increasing the open-loop gain to 50 dB generates unacceptable instability .

Results

[035] The instrument was used with 360 mL doped (7.5 mM $CuSO_4$) saline samples in bottles (o.d. = 53.5 mm, length = 187.4 mm) and a 3 T imaging magnet (Magnex, U.K.). The probe was a shielded, 79.6 mm diameter Alderman and Grant design [17] ($Q_{unloaded}$ = 310, ν_0 = 127.6 MHz) to which a small B_1 field sense coil was attached. The attenuation between the matched probe input with no sample and the sense coil output was 35 dB – the sense coil was very weakly coupled. The probe was fixed-tuned and fixed-matched for an "average" sample (20 mM NaCl, 7.5 mM Cu SO_4, Q_{loaded} = 86), with the aid of a capacitive bridge and a $\lambda/2$ balun. In-line matching (outside the magnet) was provided by a Γ-section variable filter, which has a greater dynamic range than the more usual T- or π-section filters [18]. 1 kW of class AB transmission power (PA10-90, Intech Inc. Santa Clara, CA) was available. The lines between the spectrometer and the

probe were 12 m long with a velocity factor of 0.8. In both transmission and reception, the full open-loop gain of 40 dB was realised and no instability was seen under any encountered conditions.

[036] Simple "pulse-and-acquire" experiments were performed with four samples of doped saline (0, 10, 20 and 40 mM NaCl) providing an extreme range of probe loading that would rarely be encountered in practice in a single experiment. The pH of the sample was held slightly below 7 by the addition of a very small amount of HCl to prevent precipitation of the copper. The samples were intended to act as "phantoms" for biological experiments. With the least conducting (highest Q) sample, the flip angle was set to approximately 10° (2W, 40 μs) with a repetition period of 0.9 s. A small flip angle was chosen as a "worst case" scenario, for then both transverse magnetisation and induced signal voltage were dependent on the probe Q-factor. The amplitude of the resulting FID is plotted against salt concentration in Fig. 3. As expected, there is a large change in amplitude in traditional open-loop mode, but very little change in closed-loop mode.

[037] A second experiment is shown in Fig. 4. Using a crossed diode transmit/receive switch and at low power, the transmitter output was modulated with a 1 kHz sine wave. The receiver output, which monitored the strength of the B_1 field, is shown in the figure with and without feedback, and the results speak for themselves. This facet of the Cartesian feedback technique is important for maintaining the fidelity of lower power and

selective pulses, and should also allow PIN diodes with their slow switching speeds to be dispensed with in many instances.

Conclusion

[038] In conclusion, we believe that the application of Cartesian feedback is eminently feasible for both the transmitter and receiver of an NMR spectrometer, and holds such great advantages that any instrument manufacturer that could incorporate it in a routine fashion into its product could change the manner in which NMR is practised. In operation the computer controls the set-up and calibration of the instrument, while the instrument design minimizes, and compensates for, group delay.

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Legends

[039] Figure 1. Naïve representations of negative feedback circuits for an NMR transmitter (a) and receiver (c). The on-resonance equivalents are shown respectively in (b) and (d). Oscillation is inevitable.

[040] Figure 2. Avoidance of oscillation via bandwidth reduction with the aid of Cartesian feedback. Both the transmitter and receiver are on continuously to form a feedback loop, albeit with different gains, phases and paths during transmission and reception. The usual gain-bandwidth product is 200 kHz.

[041] Figure 3. The free induction decay voltage, as a function of sample NaCl concentration, with the spectrometer in open- and closed-loop modes and subject to various manipulations. The flip angle was nominally 10^{o}. Probe loaded Q-factors are for isolated unconnected probes. For all plots, the probe was initially in-line tuned and matched for the sample with no NaCl. All voltages were measured on an oscilloscope, with an estimated error ~ ±0.5 %. The results most immune to sample conductivity were those for which the open-loop gain of the system was reset with each sample change. This task can easily be automated and rendered transparent to the user. The increase in initial FID voltage (at 0 mM) when the loop is closed was a consequence of the limited gain resolution of 1 dB (12 %). When the system is automated, the change can easily be assessed and corrected by the computer.

[042] Figure 4. The B_1 field, as monitored by the sense loop and the receiver, when a small sinusoidal voltage (e.g. a portion of a selective pulse) is applied to the transmitter modulator. Classic crossover distortion, caused by the crossed-diode transmit/receive switch, vanishes when the feedback loop is closed. Note the distortion in the waveform, indicated by the arrows, probably caused by the 1 kW power amplifier. The power spectra of the waveforms are shown on the right.

[043] A further application of Cartesian feedback is to a phased array of coils for the purpose of transmission in magnetic resonance imaging.

Introduction

[044] The use of phased-array coils for signal reception in certain specialised areas of magnetic resonance imaging is now well-established [1] and it is clear that as field strengths continue to increase, their use will become more prevalent. At least for signal reception, they yield, over elongated volumes of interest, a more homogenous spatial response function and/or improved signal-to-noise ratio (S/N). In addition, they may help to counteract propagation effects that are seen at high field strengths [2], e.g. field-focussing in head images at 8 T. Their success in so doing is, in part, because the associated mathematical manipulation of the image is a *magnitude* calculation. Naturally, it has been hoped that the same benefits would accrue with a transmit array, but there are obstacles, not least that the B_1 transmission field is the *vector and phasor* sum of the fields from individual coils and this sum is governed by Maxwell's equations. Unfortunately, at high field strengths, a volume-homogeneous B_1 field is not a solution of Maxwell's equations and a reasonably homogeneous field can only be obtained over certain surfaces [3] – e.g. a transverse slice. (However, a *magnitude*-homogeneous B_1 field may be feasible). This prohibition may easily be verified by the reader by inserting

a constant for the magnetic field **B** (i.e. a homogeneous field) in the Helmholtz equation $\nabla^2\mathbf{B}+\kappa^2\mathbf{B}=0$, where κ is the complex wavenumber. The equation can only be satisfied exactly when $\kappa = 0$; in other words, at zero frequency. In practice, the homogeneity is good enough with humans at field strengths up to 1.5 T. Notwithstanding, at higher field strengths the promise of a homogeneous B_1 field in even just a transverse slice is appealing, and array coils hold the promise of realising that appeal.

[045] In addition to the above application, transmit array coils confer in principle great flexibility in configuring specialised B_1 fields. For example, following general body imaging one might wish to examine more closely a particular region without excitation of the whole body with its high power absorption. One would then tailor powers and phases to the various coils appropriately, thereby optimising power consumption and controlling specific absorption rate (SAR). The latter mode of operation can be considered to be the counterpart of S/N optimisation with phased arrays during reception, and should be particularly advantageous at high field strengths. There are also good engineering reasons why we might wish to use array coils during transmission. When, by virtue of changing magnetic flux linkage, an induction signal is received in *any* single coil, be it a directional loop aerial for the reception of radio waves, or an NMR head coil, etc., or when that coil is used for transmission purposes, it has been realised for over half a century that as the coil's conductor length approaches a wavelength, one *must*, to retain efficiency, divide the coil into a number of smaller coils and also distribute the resultants' tuning capacitors [4] – an array *must* be created. However, this is an area fraught with difficulty,

Coil Coupling

Unloaded coils

[046] As soon as an array is created, there are couplings between the constituent coils. If we consider modern multi-conductor head coils to be tuned coils arrayed round a cylindrical surface, then couplings between coils are deliberately used to propagate energy from one array to the next. However, when one wants individual control over the currents in each coil, the essence of a phased array, the couplings are a nuisance. During reception, problems can be largely surmounted by optimal image reconstructive software

and pre-amplifier design (*q.v.*). However, the crosstalk (induced currents) [5] that results from these couplings lie at the heart of the problem of using array coils during transmission (a "Tx-Array") [6,7]. As is well known, the couplings conspire to change grossly the matching and tuning parameters from those of the isolated state; with a large number of coils, in analogy to band theory in solid state physics, a set of quasi-continuous frequency bands is formed and the interactions essentially destroy the experiment. Now as the distances between neighbouring constituent coils are usually much less than a wavelength, couplings can be considered pseudo-static in nature and classified as inductive and capacitive (magnetic and conservative-electric). Note that the use of distributed tuning capacitors helps reduce the latter coupling considerably. Such couplings have been analysed in detail since the earliest days of electronics. For example, Terman in his classic1955 textbook [4] pointed out that they could be reduced to zero in pairs of circuits. Delaying analysis of the effects of a lossy sample, as they are merely an extreme case of what follows, it is important to note that the phases of these two types of nominally-reactive couplings actually differ slightly. (We discount here 180^{o} phase differences.) The capacitive coupling phasor arising from the potentials on a coil is mainly reactive in phase when referenced to the generating current, but also has a small orthogonal resistive component, the amount depending inversely on the coil Q-factor. (We define here a resistive coupling as one that creates in a second coil a voltage in phase or anti-phase with current in a first coil. A reactive coupling is one that creates in a second coil a voltage whose phase is orthogonal to that of our current.) With substantial cancellation of the usually-dominant reactive coupling, the amplitude of the total coupling vector then becomes much smaller, but with arbitrary phase. It is not generally known that this has been understood since the earliest days of continuous wave NMR, and that a variety of still-useful cancellation schemes were used to remove this residual coupling. While today residual coupling may not be that important with pulsed (rather than continuous wave) experiments and unloaded high-Q probes, it is still important to our subject, and so we expand on the topic a little.

[047] Researchers then sought to isolate nominally-orthogonal crossed-coil probes, and to help solve the problem, resistive and inductive movable "paddles" (to steer flux), made

of metal foils having different conductivities, were used, as were various electrical bridges [8]. The paddles produced variable couplings having different amplitudes and phases, and by suitable iterative movements both the resistive and reactive components of the original coupling could be annulled. Paddles have also been used with relatively modern quadrature receiving coils [9], and variants, for use with adjacent surface coils, are shown in Fig. 5, together with a selection, by no means exhaustive, of generic and venerable bridge designs (to be inserted between the coils and replacing the two central tuning capacitors) that can cancel both resistive and reactive couplings. It is assumed that the surface coils have *negative* mutual inductance; if a design has *positive* mutual inductance (e.g. grossly overlapping surface coils or slightly overlapping crossed coils), some of the designs must be modified. Note that if the small B_1 field generated by paddles is found to be a nuisance, CR (capacitive-resistive) bridge methods may be preferable. In recent years, the reactive portions of some of these bridges have been employed [10,11], together with variable coil overlap [1], to annul reactive couplings, but as far as we know, not the resistive portions. Bridge decoupling *external* to the coils themselves was also used many years ago [12] and is the forerunner of all methods of isolation that aim to cancel induced voltages via the insertion of a decoupling impedance matrix amongst the cables from the probes [13]. It is important to note that the cancellation of *resistive* coupling entails a reduction of Q-factor, typically 1 to 5% for unloaded probes, depending on the amount of resistive coupling to be annulled. Concomitantly, there is a loss of signal-to-noise ratio (unless the extra resistance is very cold) as well as enhancement of the degree of correlation of the noise voltages in the two coils. The latter effect may easily be seen by letting voltage V_R in Fig. 5 be random. The noise current flowing in the resistive paddle then generates random equal but opposite voltages in coils 1 and 2 with this model, in addition to those created by the sample. To summarise, very accurate decoupling (-50 dB) of two coils can be achieved with care, albeit with a S/N penalty, but the process is time-consuming, unstable and in need of constant re-adjustment with change of sample or sample characteristics, temperature, humidity, etc.. For a multitude of coils, it is far too tedious and complicated and only fixed, average component values are realistic; further decoupling is really only feasible for nearest neighbours. In passing, we note that a "honeycomb" array may be helpful in

lessening the interactions with next-nearest and further neighbours when coils cover a plane, for their distance of closest approach is greater than with a square grid structure.

The loaded case

[048] Turning now to the havoc wrought by the presence of a conductive sample and in particular the human body, the latter has at radio frequencies a dominant resistive component to its frequency-dependent complex conductivity [14]. Thus eddy currents, induced by the alternating B_1 field of a first coil, generate in their turn alternating magnetic fields that induce complex voltages in all other coils. To gain insight, the effects can be simulated using one or more of Edelstein *et al*'s sample loss models [15]. For simplicity only one symmetrical model, with current and voltage phasors, is shown in Fig. 5, but several of an asymmetric nature are generally needed for accurate modelling. The model shown is a loop of conductor that couples to both coils and is predominantly resistive with some inductance. It is similar to a large resistive paddle. It therefore introduces indirect, but substantial, resistive coupling (along with a smaller reactive component) between coils. However, no new principle is present nor is a change in annulment methods needed; the annulment merely sacrifices more power on transmission and concomitantly, more S/N upon reception. The coil Q-factors are now reduced typically between 1 and 20%, depending on the degree of coupling to be cancelled. Using the model, of Fig. 5, Fig. 6 shows a simulation of the effects that sample couplings have on coil current, together with the reduction of Q-factor associated with full decoupling. The asymmetry of the coupled plot *b)* is more pronounced than we have seen in practice, due to the inadequacy of the model; nevertheless, asymmetry is often noticeable, as is the reduction of Q-factor with decoupling, so the model *is* informative.

[049] When receiving signal, a well-known way of avoiding the S/N loss associated with decoupling, and of accommodating interactions with distant neighbours, is the introduction of large, low temperature, series resistors that reduce the coil Q-factors and inhibit current flow in the coils. If the capacitive interactions and coupling due to the sample are considered to be a perturbation on the usually-dominant mutual inductive coupling, a new effective mutual induction between coils M' can be defined, together with a new coupling constant $k'_{12} = M'/\sqrt{L_1 L_2}$. L_1 and L_2 are the inductances of the two coils. Note that both M' and k'_{12} are complex and also frequency dependent, and that the *imaginary* part represents resistive loss (c.f. complex dielectric constant). The general condition for efficacy of the series resistors is then the familiar $Q << 1/|k'_{12}|$.

[050] The use of a pre-amplifier with a low input impedance for Q-reduction that effectively does this was described twenty five years ago [16,17] to speed probe ring-down, and the technique has now become a standard method of current blocking [1]. With care, the coil effective Q-factor may be reduced to the order of unity. However, this solution is of no avail during transmission. The equivalent is to mismatch the transmitter [16] so that a larger than usual resistance is effectively placed in series with the coil. Thus for a loss of 3 dB (~30%) of B_1 field strength, a resistance six times larger than usual may be inserted, but the Q-factor is reduced only by a meagre factor of seven before the power wastage becomes unacceptable. As a result, array coils are used overwhelmingly only for signal reception, while a separate, single-channel transmitter with single or quadrature coil is used for excitation. We therefore give a preliminary description of an experiment using Cartesian feedback that appears to resolve the difficulties for both transmission and

reception. We concentrate here upon the more difficult problem of driving a transmit array.

Cartesian Feedback

[051] Cartesian feedback, as it is now known in the communications industry, was first mentioned in the context of magnetic resonance in 1989 [18]. Consider *one* coil in a phased array, each coil having its own transmitter. The phased array may, with advantage, have nearest neighbour *reactive* couplings annulled by one of the methods mentioned above, as such annulment consumes no power. Our coil, as are all others, is tuned and matched for an "average" sample in the temporary absence of couplings. (All the other coils may have their tuning capacitors disconnected during the tuning and matching process.) With all coils so tuned, matched and powered, our coil's current, and the proportional B_1 field it generates, are marred by the couplings and are not that desired. Nevertheless, either by virtue of the voltage induced in a small sense loop, or by picking off a small fraction of the voltage across one of the coil's tuning capacitors, we sample that current. (The voltage method is preferable as the sense coil may also have small voltages induced by the B_1 fields from other coils. Alternatively, a very weakly coupled, shielded miniature transformer may be used.) Over a limited bandwidth that is essentially determined by the group delays in the spectrometer, that current sample is then amplified in the spectrometer's receiver, phase-sensitively detected in quadrature with the appropriate phase, and compared with the value it is supposed to be, as determined by the complex voltage applied to the transmitter quadrature modulator. Any erroneous difference in the two is then applied as a correction in the usual feedback manner familiar to those who use operational amplifiers. This is the quintessential negative feedback technique of audio amplification applied successfully in quadrature at radio frequencies by severely restricting, in the audio section of the transmitter, the bandwidth over which feedback pertains.

[052] It is shown above that the feedback effectively makes the transmitter a constant current generator *with no change of power efficiency*, and the greater the loop gain of the

feedback system, the more closely this state is approximated. Thus the now-correct current and its B_1 field become impervious to changes of sample loading, while voltages induced in our coil from other coils, directly or via the intermediary of a sample, have negligible effect on the current flowing – the crosstalk has substantially been eliminated. If the transmitter, after the impedance transformation associated with matching, is considered to be in series with our tuned surface coil, with output impedance initially equalling the effective resistance of that coil when isolated, it can be shown that the feedback increases that output impedance by the order of the open-loop gain. Thus the Q-factor of the coil would be diminished by the same amount – say a factor of 100 – if the effect were broadband. As we move off resonance, however, the loop gain decreases, thanks to the limited bandwidth, and so the beneficial effects of the feedback also decrease. Of prime importance then is the bandwidth over which the full loop gain can be applied. We currently can safely use in our Cartesian feedback instrument (no risk of oscillation) a gain bandwidth product of 200 kHz centred on the Larmor frequency, and so typically apply 40 dB of loop gain using audio RC filters that have ±1 kHz bandwidth. Thus the full benefits of the feedback should be available over a 2 kHz bandwidth about the Larmor frequency, and this should be compared with the bandwidth of a typical selective pulse – they are about the same. Notwithstanding our having invested a large amount of time and effort to obtain this gain-bandwidth product, it would be advantageous to increase it, and we are now aggressively exploring the use of negative group delay techniques to achieve this goal.

Experimental

[053] To test the validity of the above idea, we have at present only one 128 MHz feedback spectrometer, and so have had to tailor an experiment accordingly. Thus, only the two-coil array of Fig.5 was employed, modified as shown in Fig. 7. When individually sitting above a large spinal imaging saline phantom, both square coils were tuned and nominally matched to 50 Ω with the aid of a half wavelength balun at 126.6 MHz – a convenient frequency determined by available capacitors and within the tuning range of the spectrometer. Unloaded, the Q-factor of each coil was ~420; loaded the factors dramatically reduced to 16 and 15. One coil was slightly further from the phantom

(7.2 mm) than the other (4.3 mm) so that the coils could be overlapped if desired. The coils were then brought together as shown, and with the aid of a network analyser driving the second coil and receiving signal from the first coil (current monitors not yet used), the coil coupling was measured: it was –9.4 dB. The reactive coupling between the coils was now annulled with a tuned reactive paddle that resonated at 135.5 MHz with a Q-factor of 304 and a tuning capacitor of 7.5 pF. Thus the impedance of the paddle was 0.6 - 21j Ω – not a particularly pure reactive condition, but adequate for our purposes. The coupling between the two coils' ports was, however, only reduced to –13.8 dB, showing that considerable resistive coupling existed via the sample. The input impedance of each coil was found to have dropped by roughly 2 Ω.

[054] With the network analyser still driving the second coil, the first coil was now connected to the quiescent but operational transmitter of the Cartesian feedback spectrometer. For the safety of the network analyser, lest anything should go wrong, a very low power (1W maximum) amplifier of nominal output impedance 50 Ω was employed at the end of the transmitter chain. Two balanced current sensors on the first coil were now made, as shown, by lightly tapping the voltages across two tuning capacitors. The first tap was connected to the spectrometer's receiver, but with the feedback loop open for the moment. In other words, there was normal transmitter operation but with no voltage applied to the transmitter modulator - the transmitter was merely functioning as a nominal 50 Ω load on the first coil's input. The second current-monitoring tap was then attached to the receive port of the network analyser and a *relative* measure of the on-resonance current flowing in the first coil was taken – it was – 48.6 dB. From the values of the tap capacitors (0.2 pF), we might have expected –50.5 dB, corresponding to the previously-measured –13.8 dB coupling, but both the values of the tapping capacitors and the output resistance of the spectrometer's transmitter were quite nominal. The spectrometer feedback, with 1 kHz filters and nominally 40 dB open-loop gain, was now turned on. The reduction in the current in the first coil was dramatic (-40.4 dB), as shown in Fig 8a, and is in reasonable agreement with the simplistic theory, found in any elementary electronics text, that the current amplitude I should vary as

$$I = \left| \frac{I_0}{1 + \alpha F(\delta\omega)} \right| \tag{1}$$

where α is the open-loop gain (100) and $F(\delta\omega)$ is the single pole, low-pass filter response function

$$F(\delta\omega) = (1 + j\delta\omega / \omega_c)^{-1} \tag{2}$$

Here, $\delta\omega$ is the distance off resonance, positive or negative, and ω_c is the filter cut-off frequency (1 kHz). As expected, the reduction diminishes off resonance. The experiment was now repeated with the decoupling paddle removed so that there was reactive as well as resistive coupling between the coils. The induced current, not surprisingly, increases in the absence of feedback, Fig. 8b, but in the presence of feedback, the change in the shape of the response is noteworthy, and can only be explained with a full simulation (c.f. the asymmetry of Fig. 6b). The basic efficacy of the technique, however, remained the same, and importantly, there was no sign of current "peaking" – a prelude to possible oscillation. Finally, the experiment was repeated with the mutual induction's being cancelled by overlapping the coils. The results are shown in Fig. 7 and are similar.

[055] A point of concern was whether the transmitter on the first coil would have to try and exceed its maximum output voltage rating, to do the job of opposing induced voltages, if a similar transmitter were attached to the second coil and turned fully on. It may be shown that the overload criterion for a coil-pair with matched transmitters is $|k_{12}| < 2/Q$, where k_{12} is the effective complex coupling factor between the two coils defined earlier. We therefore measured k_{12} and found it to be $-0.0068 + 0.072j$. With a Q-factor of 16, the criterion was therefore just met; however, with cancellation of reactive coupling it was easily met, and as the real part of k_{12} and 1/Q ride hand in hand, we would expect this generally to be so. In this regard, it is worth remembering that *in extremis*, a mismatch of the transmitters is available to provide assistance, as is resistive decoupling at the expense of some power and signal-to-noise ratio.

Conclusion

[056] These experiments demonstrate that it is possible using Cartesian feedback to drive normally and efficiently a first tuned, matched and heavily sample-loaded coil in the presence of a second coupled coil in close proximity. Cartesian feedback applied to the transmitter attached to that second coil effectively introduces therein a high series resistance, which inhibits current from flowing and prevents a back-EMF from being induced in the first coil – the crosstalk is blocked. For optimal use, any large reactive coupling between coils should first be annulled with one of the various bridge methods available. The technique can make a multi-Tx phased array practical, each element being driven from its own Cartesian feedback instrument. It is stressed that no extra transmitter power is needed to produce a given B_1 field when feedback is invoked. This, in turn, opens the door to all the projected advantages of array coils for transmission – localisation of B_1 fields, control of SAR, production of B_1 fields have a specific spatial variation, and the creation of homogeneous B_1 fields at high frequencies over those surfaces where it is theoretically possible.

[057] While we have concentrated on transmission, it must be emphasised that it is to be expected that similar effects hold for signal reception, and Cartesian feedback can generally provide better decoupling with no loss of S/N than can pre-amplifier damping, though over a smaller bandwidth. Further, there is no reason both techniques should not be employed simultaneously. The Cartesian feedback technique is neither simple nor cheap, but the cost of multiple transmitters is offset by the fact that individual power amplifiers need to generate much less power than a single main unit, and so are concomitantly less expensive than the latter. The total expense is then about the same. While the available bandwidth is comparable with that of a typical selective pulse.

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Legends

[058] Figure 5. In a), a pair of surface coils 1 and 2 with a lossy sample model (S) and adaptations of paddles for reactive (X) and resistive (R) decoupling. Conventions: induced EMFs obey Lenz's Law and so follow a *left* hand rule with reference to the flux linkage that creates them; current directions are always shown so that their phasors are in the first or fourth quadrants. When the surface coils are separated as shown, both reactive components of the induced voltages in coil 2 enhance; the resistive component from the sample is negative. Typical measured values for dimensions ~ 10 cm and frequencies ~ 128 MHz are $V_{S2} + V_{12} \sim (-1 + j10)\, i_1$. The induced currents that flow in the two

corrective paddles are nominally in phase quadrature, and by appropriate adjustment of the paddle positions and/or geometry, couplings of arbitrary amplitude and phase can be annulled. Note that the additional B_1 fields generated by paddles may be unacceptable. Figure 5b) is a combined resistive and reactive paddle; in some circumstances an additional capacitor C_3 may be needed. Diagrams c) through f) are examples of decoupling bridges to be inserted between the two coils, replacing the two central tuning capacitors. Maintenance of those capacitors' reactances, in the form of the input reactances of the chosen bridge, is usually possible. It is assumed that the annulling EMF the bridges must transfer is in the fourth quadrant; viz $(a - jb)$. If this is not so, design changes will be needed. The transformer-based bridges c) and d) are included for generality and are not recommended: winding capacitance, uncertain mutual inductance and stray couplings to the surface coils render them difficult to use. e) is an RC bridge of great flexibility as both C_2 and/or R_2 can be parallel pairs or crossed pairs. e) is the simplest bridge as only C_2 and R_2 determine the coupling while C_1 is used for reactance maintenance. It is usually applicable to all nearest neighbour, co-planar separated coils.

[059] Figure 6. A simulation of coupling effects using a single Edelstein sample model as per Fig. 5. The current in driven coil 1 is shown *a*) with coil 2 open circuit; *b*) with direct mutually inductive and indirect sample coupling to coil 2 – note the asymmetry (not as pronounced in practice), in marked contrast to the usual portrayal of coupling that splits the resonance, and *c*) with both resistive and reactive decoupling applied with the decoupling network of Fig. 5f – note the decreased Q-factor. The curves correspond roughly to those observed regardless of the decoupling method. (N.B. both coils were matched to 50 Ω hence the broadness of the curves.)

[060] Figure 7. The equipment used to demonstrate current reduction when Cartesian feedback is applied to a transmitter. Above a saline phantom, surface coil 2 is driven by a network analyser that also records the induced current flowing in coil 1 as a result of coupling. Coil 1 is connected to an active but quiescent transmitter equipped with Cartesian feedback. Reactive decoupling is provided by a paddle. A second measure of

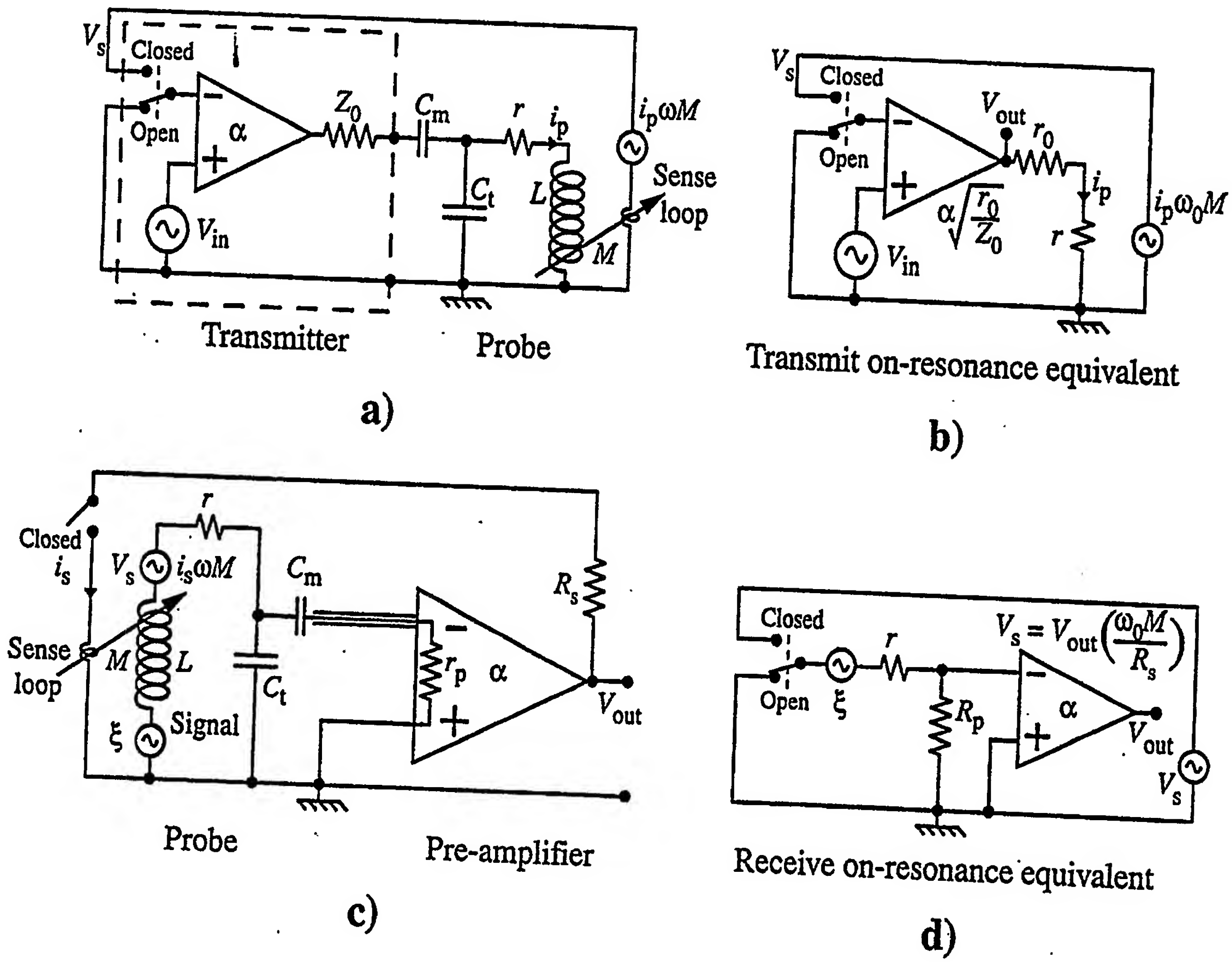
the current in coil 1 is provided to the receiver of the spectrometer and is used to generate a voltage from the transmitter that greatly reduces the current.

[061] Figure 8. Induced current in coil 1 of figure 7 when coil 2 is driven in the vicinity of resonance by the network analyser. Coil 1 is connected to a Cartesian feedback transmitter. In a), reactive coupling between the coils was cancelled with the aid of a paddle. In b), the paddle was removed – c.f. the asymmetry in Fig.6b. The spectrometer's and the network analyser's frequency sources disagreed by 4.6 kHz (36 ppm), hence the plots' shifts to the left.

[062] Figure 9. The two coils of figure 6 were overlapped to remove reactive coupling. In a), the overlap was adjusted for zero mutually inductive coupling in the *absence* of the phantom. However, in the *presence* of the phantom, the shift and depression of the experimental feedback current curve indicated that reactive coupling was still present, presumably contributed by the sample. In b) the total coupling was minimised by changing the overlap of the coils. This presumably removed all reactive coupling, thereby generating a current curve with the expected dip and negligible frequency shift.

Abstract

[063] The use of Cartesian electronic feedback for effecting a major improvement in the functioning of magnetic resonance instrumentation such as the transmitter and receiver of an NMR spectrometer is disclosed. The dependences of both pulse power/length and signal strength upon probe loading, matching and tuning are virtually eliminated. Thus, for a given probe, sample geometry and flip angle, the free induction decay signal strength is rendered solely dependent upon the number of nuclei. The instrument therefore becomes capable of absolute calibration. In addition, phase and amplitude distortion of selective pulses, introduced by crossed diodes, power amplifier heating, etc., is virtually eliminated, as are radiation damping and phase modulation caused by probe vibration. The use of multiple probes at the same frequency, for example quadrature probes and phased arrays, is also simplified as the interactions between such probes are typically reduced by two orders of magnitude. The application of Cartesian electronic feedback to the use of a phased array of coils for the purposes of transmission in magnetic resonance imaging is also disclosed.



Vs
Closed
Open
α
Vin
Z0
Cm
r
ip
Ct
L
ipωM
Sense loop
M
Transmitter
Probe
a)
Vout
r0
α√(r0/Z0)
ipω0M
Transmit on-resonance equivalent
b)
is
isωM
Signal
ξ
rp
Rs
Vout
Probe
Pre-amplifier
c)
Rp
Vs = Vout(ω0M/Rs)
Receive on-resonance equivalent
d)

Figure 1

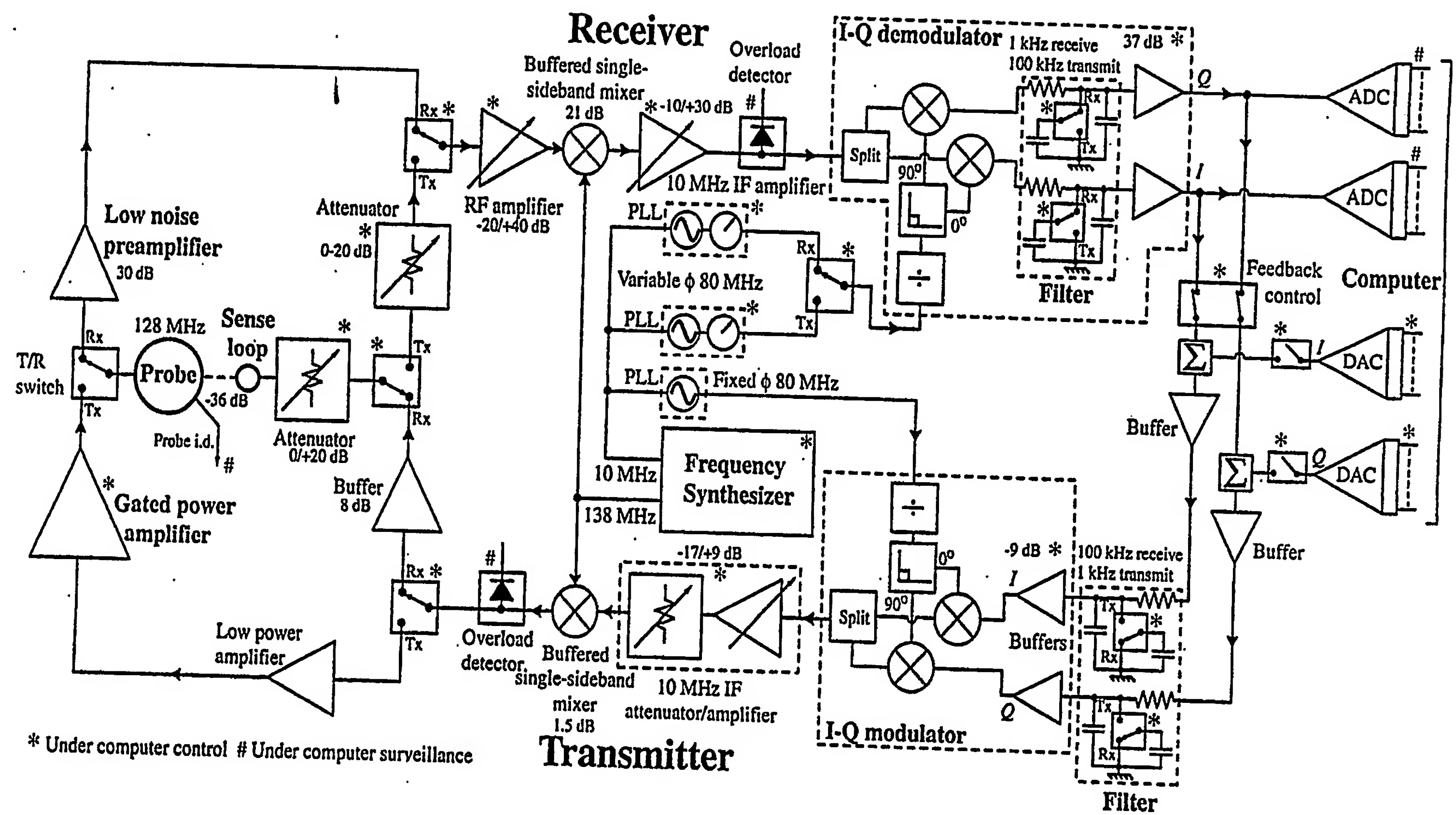

Receiver
Transmitter
Low noise preamplifier 30 dB
T/R switch
Rx
Tx
Probe
128 MHz
Sense loop
-36 dB
Probe i.d.
Gated power amplifier
Attenuator 0-20 dB
Attenuator 0/+20 dB
Buffer 8 dB
Low power amplifier
RF amplifier -20/+40 dB
Buffered single-sideband mixer 21 dB
-10/+30 dB
10 MHz IF amplifier
Overload detector
PLL
Variable ϕ 80 MHz
Fixed ϕ 80 MHz
Frequency Synthesizer
10 MHz
138 MHz
Overload detector
Buffered single-sideband mixer 1.5 dB
-17/+9 dB
10 MHz IF attenuator/amplifier
I-Q demodulator
I-Q modulator
Split
90°
0°
1 kHz receive
100 kHz transmit
100 kHz receive
1 kHz transmit
37 dB
-9 dB
Filter
Buffers
Buffer
Feedback control
Computer
ADC
DAC
I
Q
* Under computer control # Under computer surveillance

Figure 2

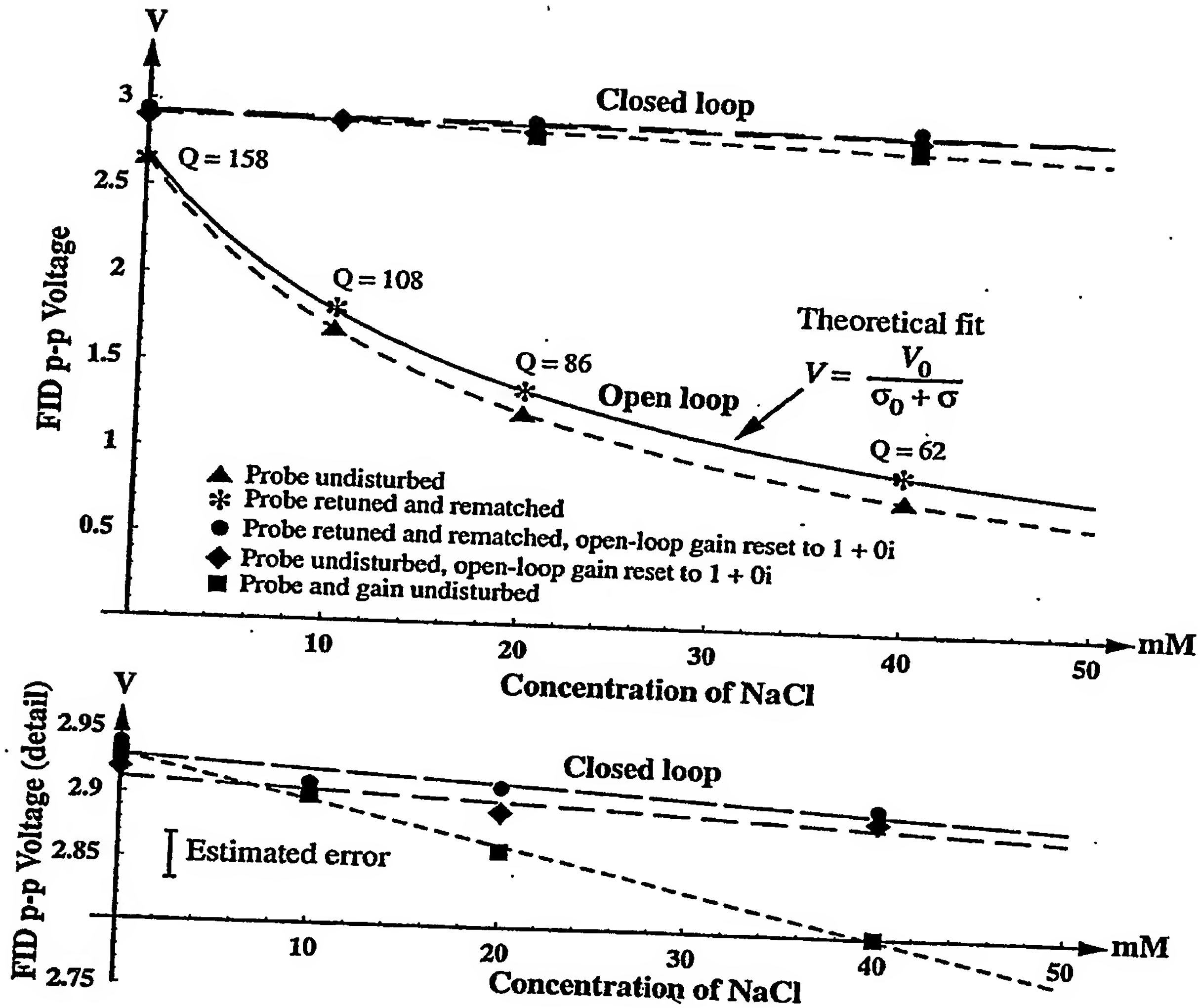

V
Closed loop
Q = 158
Q = 108
Q = 86
Q = 62
Open loop
Theoretical fit
$V = \frac{V_0}{\sigma_0 + \sigma}$
FID p-p Voltage
3
2.5
2
1.5
1
0.5
Probe undisturbed
Probe retuned and rematched
Probe retuned and rematched, open-loop gain reset to 1 + 0i
Probe undisturbed, open-loop gain reset to 1 + 0i
Probe and gain undisturbed
10
20
30
40
50
mM
Concentration of NaCl
V
FID p-p Voltage (detail)
2.95
2.9
2.85
2.75
Closed loop
Estimated error
10
20
30
40
50
mM
Concentration of NaCl

Figure 3

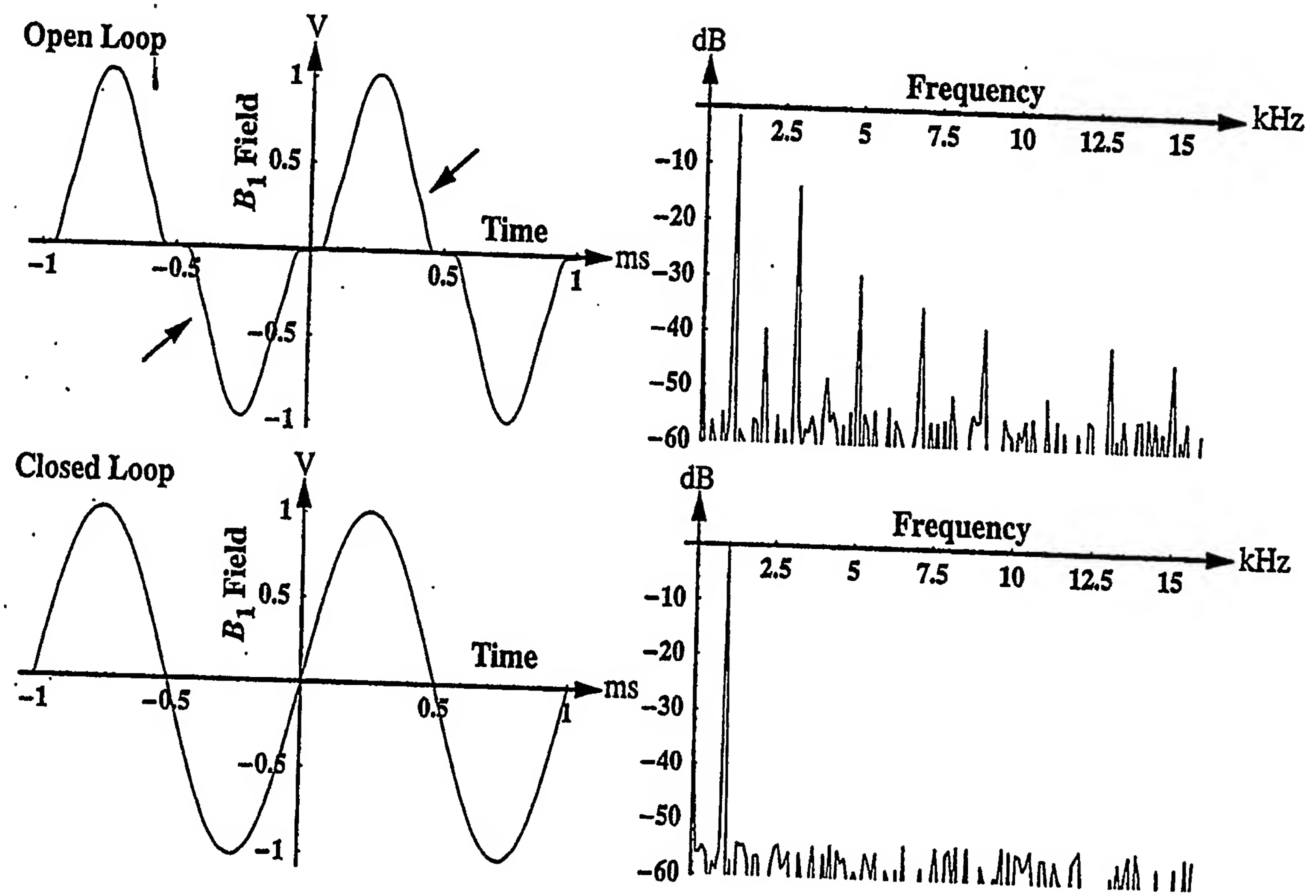
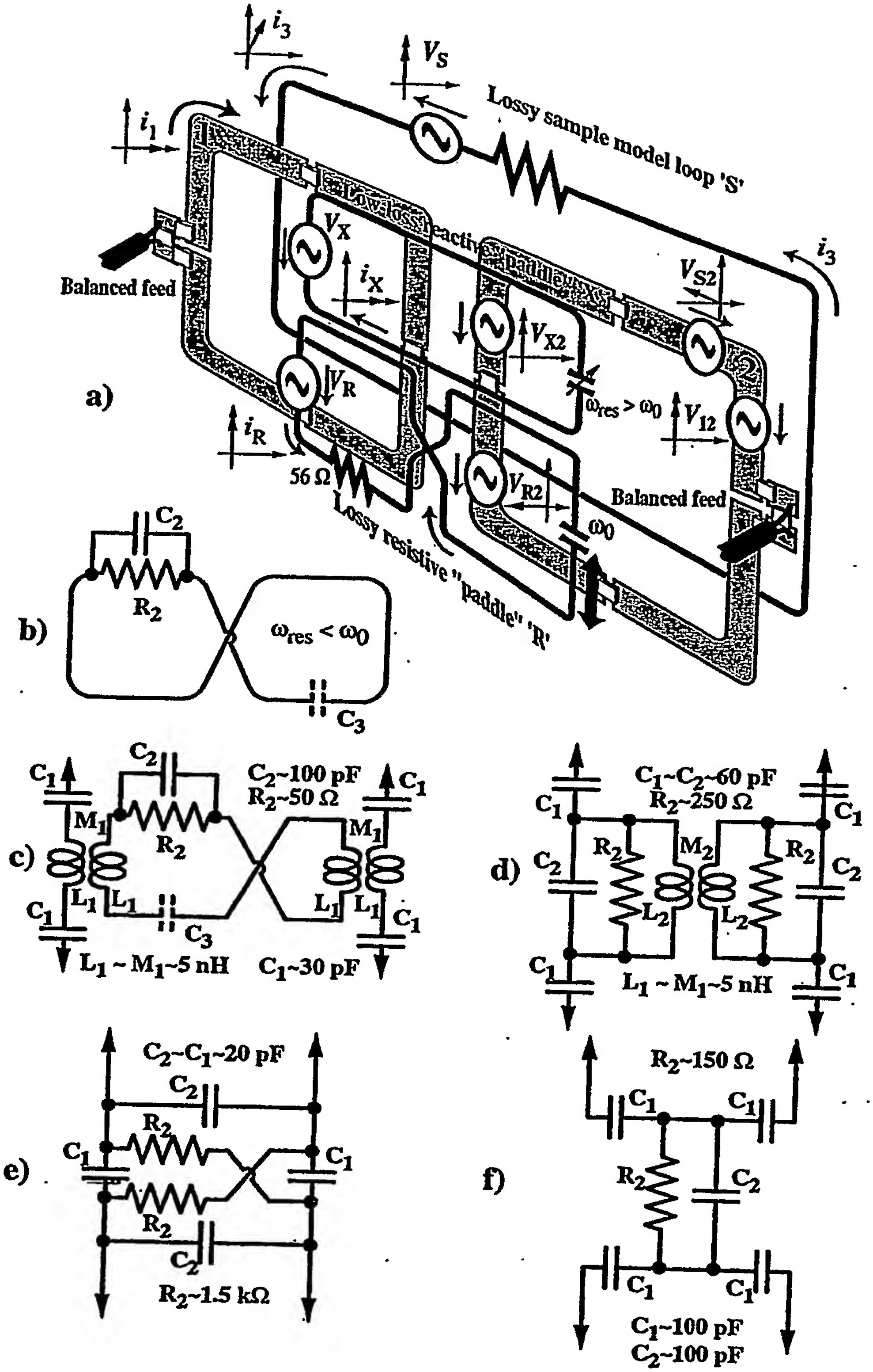
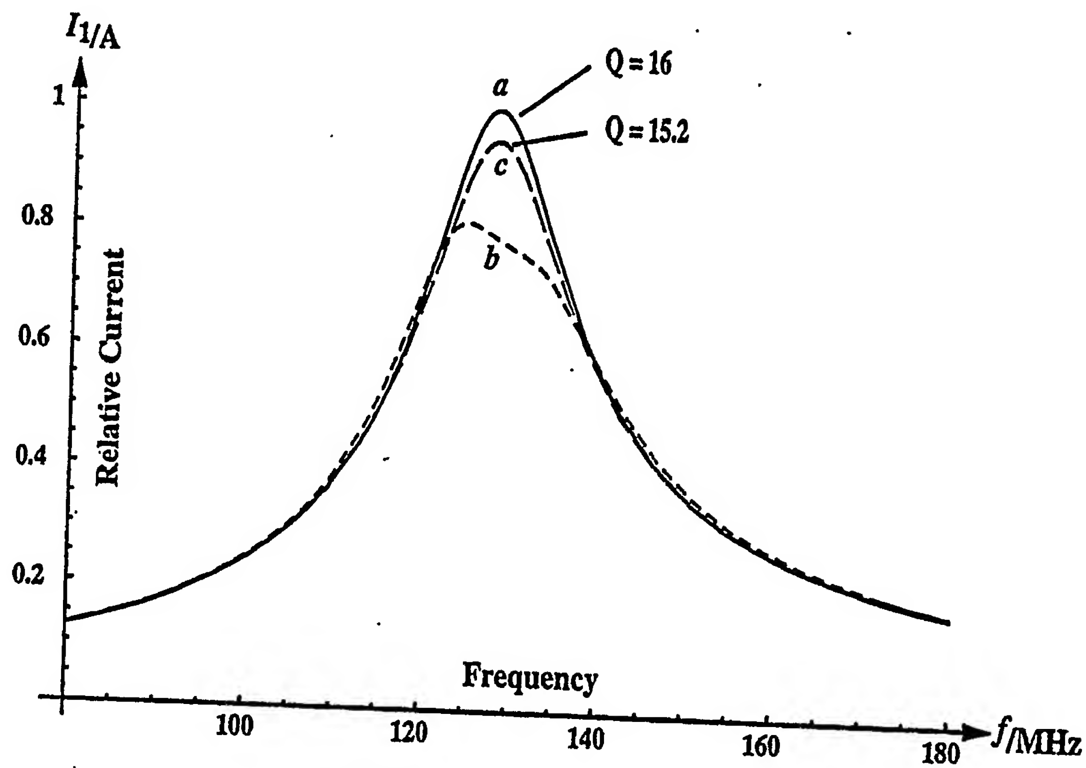

Open Loop
Closed Loop
V
B_1 Field
Time
ms
1
0.5
-0.5
-1
dB
Frequency
kHz
2.5
5
7.5
10
12.5
15
-10
-20
-30
-40
-50
-60

Figure 4



i_3
V_S
i_1
Lossy sample model loop 'S'
Low-loss reactive "paddle" 'X'
Balanced feed
V_X
i_X
V_{S2}
i_3
V_{X2}
V_R
a)
$\omega_{res} > \omega_0$
V_{12}
i_R
56 Ω
V_{R2}
Balanced feed
Lossy resistive "paddle" 'R'
ω_0
b)
C_2
R_2
$\omega_{res} < \omega_0$
C_3
c)
C_2~100 pF
R_2~50 Ω
L_1 ~ M_1~5 nH
C_1~30 pF
d)
C_1~C_2~60 pF
R_2~250 Ω
L_1 ~ M_1~5 nH
e)
C_2~C_1~20 pF
R_2~1.5 kΩ
f)
R_2~150 Ω
C_1~100 pF
C_2~100 pF

Figure 5



$I_{1/A}$
1
0.8
0.6
0.4
0.2
Relative Current
a
Q = 16
Q = 15.2
c
b
Frequency
100
120
140
160
180
$f_{/MHz}$

Figure 6

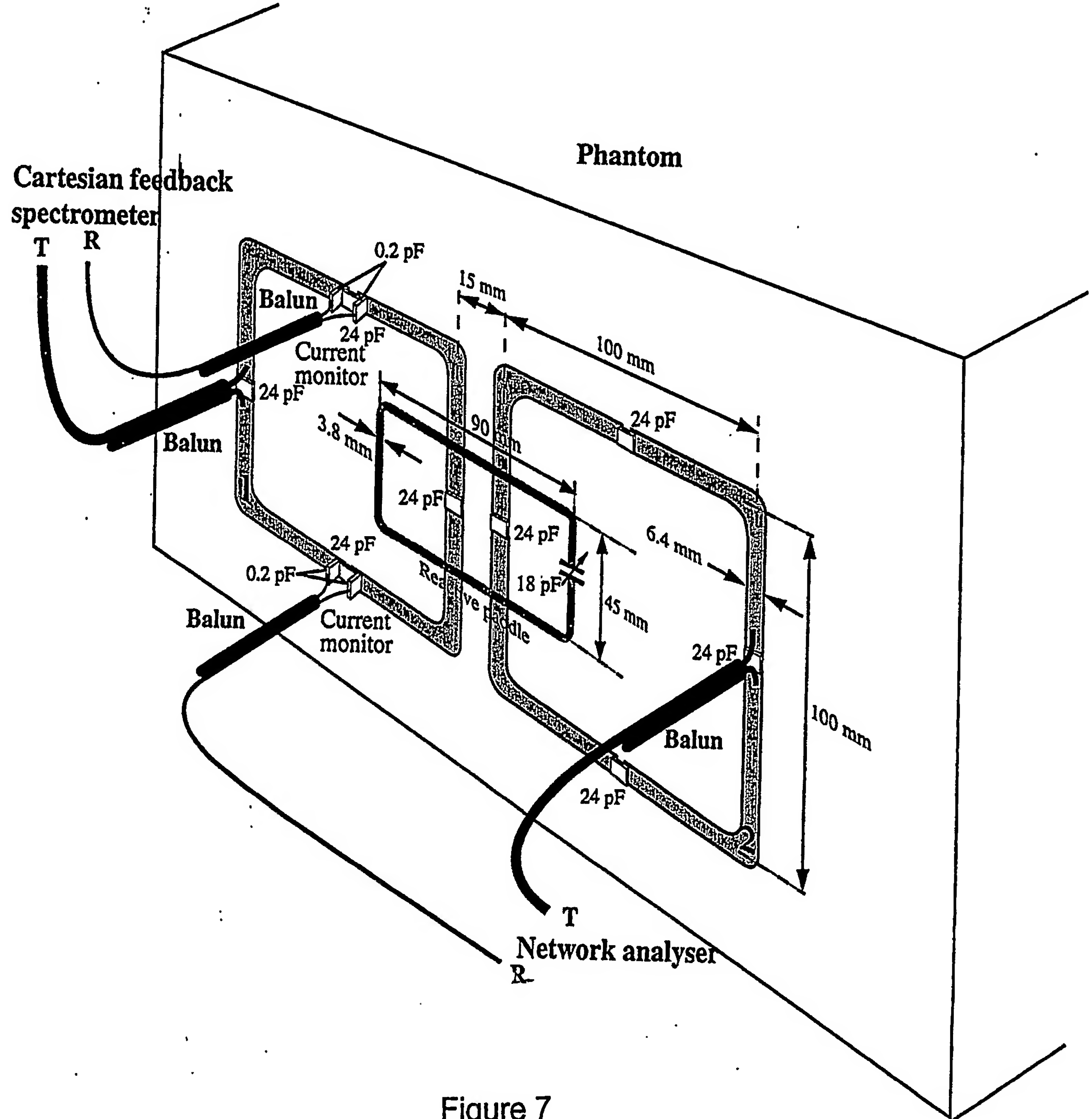
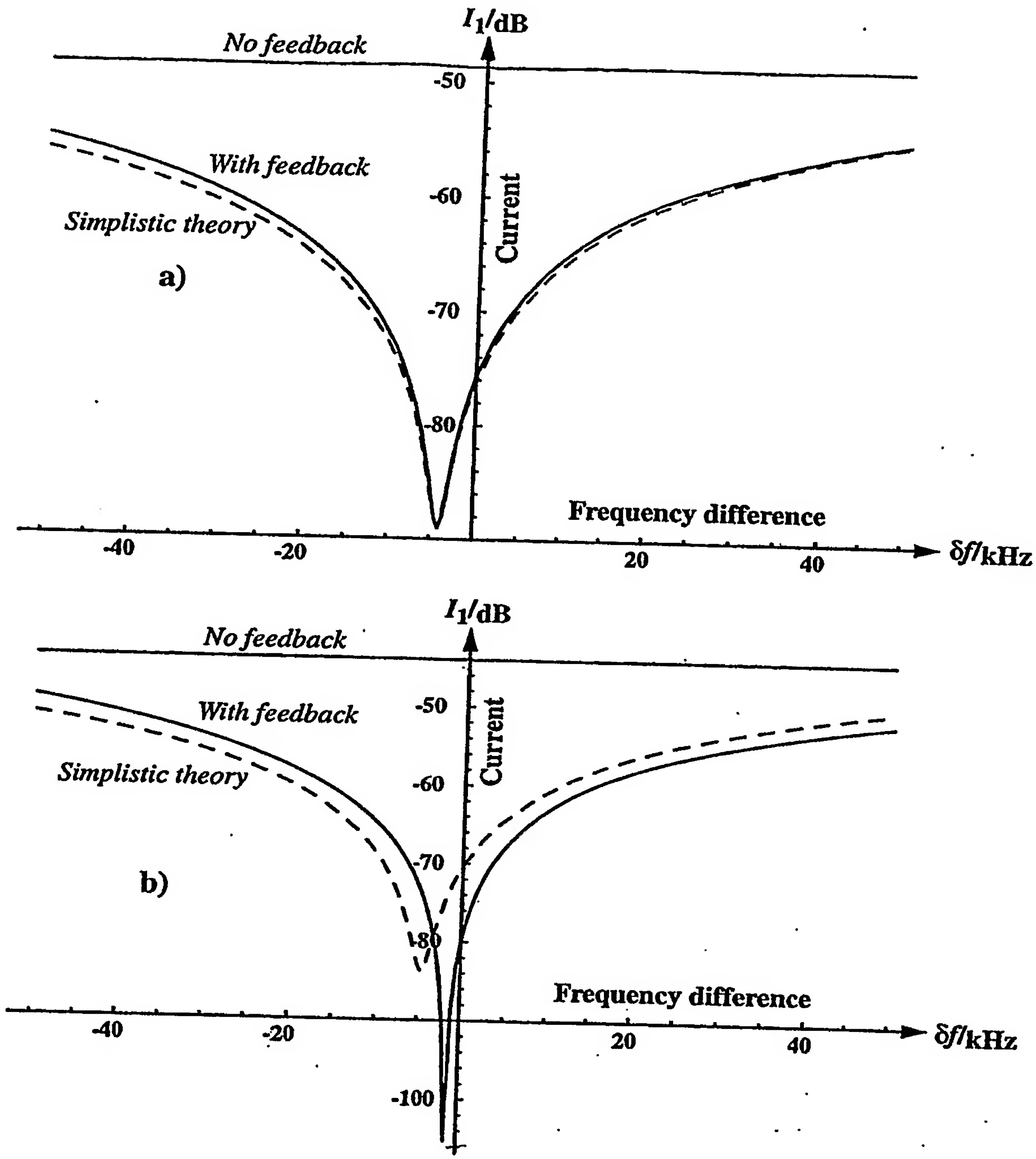
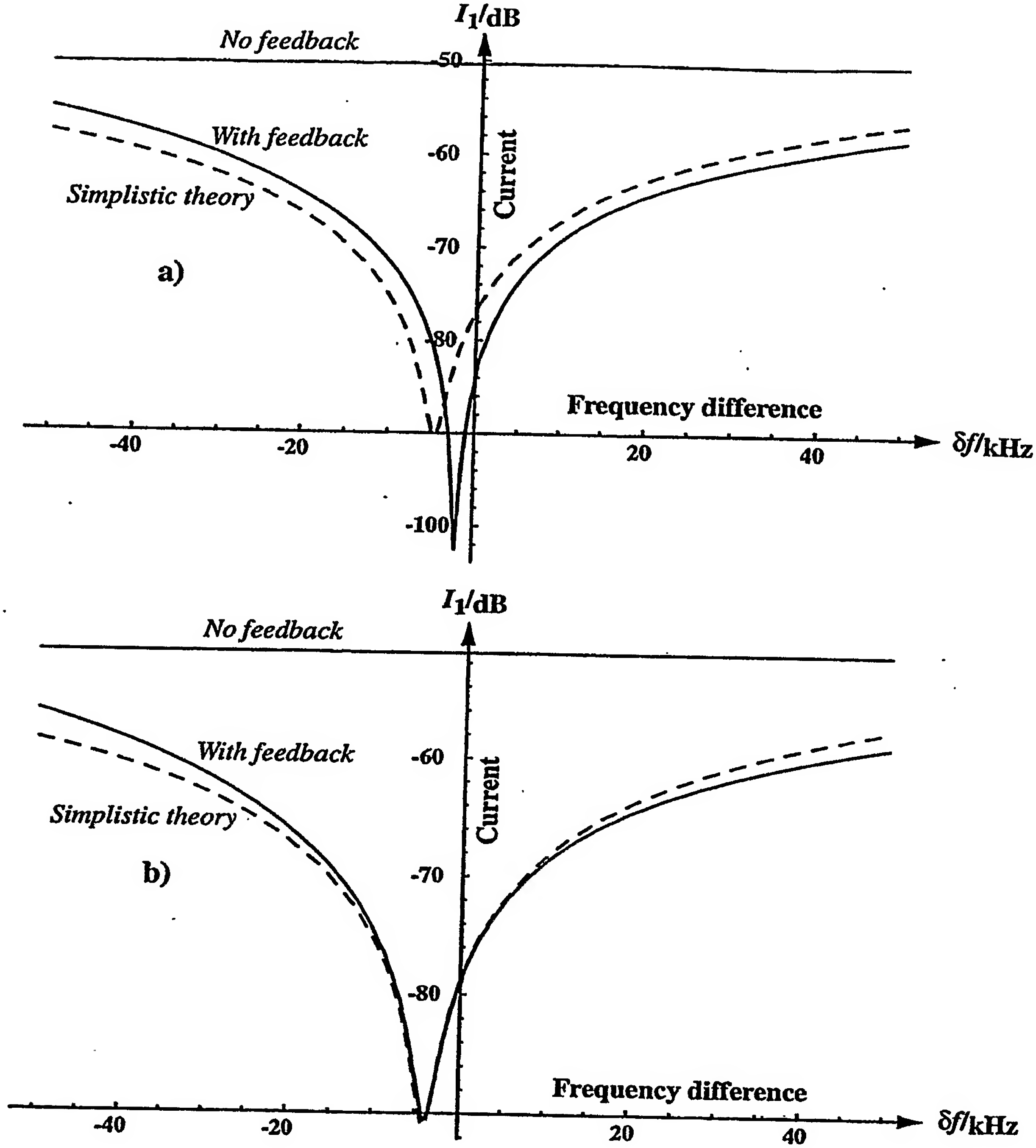

Phantom
Cartesian feedback spectrometer
T
R
Balun
0.2 pF
24 pF
Current monitor
24 pF
Balun
3.8 mm
15 mm
100 mm
24 pF
90
24 pF
24 pF
6.4 mm
18 pF
45 mm
24 pF
0.2 pF
Balun
Current monitor
24 pF
Balun
24 pF
100 mm
T
Network analyser
R

Figure 7



a)
I_1/dB
No feedback
With feedback
Simplistic theory
Current
-50
-60
-70
-80
Frequency difference
-40
-20
20
40
δf/kHz
b)
I_1/dB
No feedback
With feedback
Simplistic theory
Current
-50
-60
-70
-80
-100
Frequency difference
-40
-20
20
40
δf/kHz

Figure 8



I_1/dB
No feedback
-50
With feedback
-60
Simplistic theory
Current
-70
a)
-80
Frequency difference
-40
-20
20
40
δf/kHz
-100
I_1/dB
No feedback
With feedback
-60
Simplistic theory
Current
b)
-70
-80
Frequency difference
-40
-20
20
40
δf/kHz

Figure 9